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TITLE OF THE INVENTION

ACTIVE RESISTIVE SUMMER FOR A
TRANSFORMER HYBRID

BACKGROUND OF THE INVENTION

Field of the Invention

The present invention relates generally to transmitting and

receiving electrical signals through communication

channels, such as a gigabit channel. In particular, the

present invention relates to a transmit canceller that

removes transmit signals from receive signals in such

communication channels.

Background and Related Art

A gigabit channel is a communications channel with a total

data throughput of one gigabit per second. A gigabit

channel typically includes four (4) unshielded twisted

pairs (hereinafter "UTP") of cables (e.g., category 5

cables) to achieve this data rate. IEEE Standard 802.3ab,

herein incorporated by reference, specifies the physical

layer parameters for a 1000BASE-T channel (e.g., a gigabit

channel).

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As will be appreciated by those skilled in the art, a UTP becomes a transmission line when transmitting high frequency signals. A transmission line can be modeled as a network of inductors, capacitors and resistors, as shown in

5   Figure 1. With reference to Figure 1, G is normally zero and $R(\omega)$ is complex due to skin effect. $R(\omega)$ can be defined by:

$$R(\omega) = k_R(1 + j)\sqrt{\omega}, \qquad (1)$$

where k_R is a function of the conductor diameter,

10   permeability, and conductivity. The characteristic impedance of the line is defined by:

$$Z_0 = \sqrt{\frac{R(\omega) + j\omega L}{G + j\omega C}}, \qquad (2)$$

and at high frequencies, Z_0 becomes approximately $\sqrt{L/C}$ or approximately 100 ohms in a typical configuration. When

15   properly terminated, a UTP of length d has a transfer function H that is a function of both length (d) and fre-quency (ω):

$$H(d, \omega) = e^{d\gamma(\omega)}, \qquad (3)$$

20   where

$$\gamma\omega = \sqrt{(R(\omega) + j\omega L)(G + j\omega C)}, \qquad (4)$$

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and substituting Equations 1 and 4 into Equation 3, and simplifying, approximately yields:

$$H(d,\omega) \approx \exp\left\{d\left[\frac{k_\iota}{2}\sqrt{\frac{\omega L}{C}} + j\left(\omega\sqrt{LC} + \frac{k_\iota}{2}\sqrt{\frac{\omega L}{C}}\right)\right]\right\}. \qquad (5)$$

5   Equation 5 shows that attenuation and delay are a function of the cable length d.

A transmission path for a UTP typically includes a twisted pair of cables that are coupled to transformers at both a
10 near and far end, as shown in Figure 2. A transceiver at each end of the transmission path transmits and receives via the same twisted pair. A cable typically includes two patch cords totaling less than 10m, and a main section of 100m or even longer. The transmitters shown in Figure 2
15 are modeled as current sources. The near end current source supplies a current I_{tx}. The near end transmit voltage (e.g., $I_{tx}R_{tx}$) is detected and measured across resistor R_{tx}. A receive signal V_{rcv} (e.g., a signal transmitted from the far-end transceiver) is also detected
20 and measured across resistor R_{tx}. Hence, V_{tx} includes both transmit ($I_{tx}R_{tx}$) and receive (V_{rcv}) signals. Accordingly, the signal V_{rcv} (e.g., the signal from Transceiver B)

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received at Transceiver A can be obtained by taking the

difference between the transmit voltage and the measured

voltage V_{tx}, as follows:

5                $$V_{rcv} = V_{tx} - I_{tx}R_{tx}. \tag{6}$$

Conventional solutions for removing transmit signals from

receive signals often employ known transconductor ("Gm")

10   summing stages or other current based methods. As will be

appreciated, these methods often introduce signal

distortion into the receive signal. Also, some

transconductors have a limited signal dynamic range.

Accordingly, conventional methods are often inadequate for

15   applications requiring signal recovery. Additionally, known

summing circuits, such as weighted summers using

operational amplifiers, have not heretofore been modified

to accommodate the intricacies associated with canceling

transmit signals or regulating baseline wander (described

20   below). A known weighted summer is discussed in Chapter 2

of "Microelectronic Circuits, Third Edition," by A.S. Sedra

and K.C. Smith, 1991, incorporated herein by reference.

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As will be appreciated by those skilled in the art, the
receive signal V_{rcv} typically contains additional components,
due to baseline wander, echoes and crosstalk, for example.

5   Baseline wander is preferably corrected for when
transmitting and receiving signals over transmission lines.
Removing DC components from a receive signal using
transformer coupling can cause baseline wander. As will be
appreciated by those skilled in the art, baseline wander
10   represents a deviation from an initial DC potential of a
signal.

"Echoes" typically represent a residual transmit signal
caused by reflections that appear in the receive signal.
15   Echoes can cause undue interference depending on the size
of the reflection.

Capacitive coupling between the channels, as shown in
Figure 3, causes crosstalk. Four channels TX1-TX4 are
20   shown in Figure 3. The capacitive coupling between TX1 and
each of TX2, TX3 and TX4 are modeled by capacitors C_{1-2}, C_{1-3},
C_{1-4}, respectively. The capacitive coupling forms a high-
pass filter between channels and therefore crosstalk con-
tains mostly high frequency components. As will be

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appreciated by those skilled in the art, normally only the

near-end crosstalk (NEXT) needs to be considered, since

crosstalk is usually small and the transmission line

provides further attenuation of the far-end crosstalk

5   (FEXT).

Accordingly, there are many signal-to-noise problems to be

solved in the art. Hence, an efficient transmission

canceller is needed to remove a transmit signal from a

10   receive signal without introducing excess signal

distortion. An electrical circuit is also needed to

subtract a transmit signal from a receive signal. There is

a further need of an electrical circuit to correct baseline

wander.

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SUMMARY OF THE INVENTION

The present invention relates to a transmit signal

canceller for use in a transformer hybrid. Such a hybrid

includes a junction for transmitting and receiving signals.

20   In the present invention, an active resistive summer can be

used to cancel a transmit signal from a receive signal.

According to the invention, an electrical circuit in a communications channel is provided. The electrical circuit includes an active resistive summer having: (i) an input for a composite signal, the composite signal including a

5   transmission signal component and a receive signal component, (ii) an input for a replica transmission signal, and (iii) an output for a receive signal which includes the composite signal minus the replica signal.

10   According to an another aspect of the present invention, a transmit signal canceller in a communication channel is provided. The channel includes a first transceiver for transmitting and receiving signals and a replica transmitter for generating a replica transmission signal

15   input. A composite signal at a rear end includes a transmission signal of the first transceiver and a received signal of a second transceiver. The transmit canceller includes: (i) an operational amplifier having a positive input terminal, a negative input terminal, and an output

20   terminal; (ii) a feedback element in communication with the negative input terminal and the output terminal; (iii) a first input resistor in communication with the negative input terminal and the measured signal input; (iv) a second input resistor in communication with the negative input

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terminal and the replica signal input; and (v) a
predetermined voltage source in communication with the
positive terminal of the operational amplifier. The
receive signal is an output at the output terminal of the
5   operational amplifier.

According to still another aspect of the present invention,
a communication system including a first transmission
channel with a first end and a second end is provided. The
10 first end couples to a first transformer and the second end
couples to a second transformer. A first transceiver
transmits and receives signals via the first transformer
and a second transceiver transmits and receives signals via
the second transformer. A first signal is supplied at the
15 near end. The first signal includes a transmission signal
component of the first transceiver and a receive signal
component of the second transceiver. The communications
system includes: (i) a replica transmitter that generates a
replica of the transmission signal component of the first
20 transceiver; (ii) a filter to filter the replica signal;
(iii) an active resistive summer receiving the first
signal, and the filtered replica signal as inputs to reduce
the transmission signal component at an output of the
active resistive summer.

According to still another aspect of the present invention, a method of correcting baseline wander in a receive signal in a communications channel having a near and far end is

5   provided. The channel includes a first transceiver at the near end and a second transceiver at the far end, each to transmit and receive signals. The method includes the steps of: (i) providing a composite signal, the composite signal including a transmission signal of the first

10   transceiver and a receive signal of the second transceiver; (ii) generating a replica of the transmission signal; (iii) subtracting the replica signal from the composite signal through an active resistive summer; and (iv) providing a baseline correction current into the active resistive

15   summer.

These and other objects, features, and advantages of the present invention will be apparent from the following description of the preferred embodiments of the present

20   invention.

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BRIEF DESCRIPTION OF THE DRAWINGS

The details of the present invention will be more readily understood from a detailed description of the preferred

5   embodiments taken in conjunction with the following figures.

Figure 1 is a circuit diagram illustrating a transmission line model.

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Figure 2 is a circuit diagram illustrating a transmission path across a twisted pair of cables, the cables being coupled to transformers at each end.

15   Figure 3 is a diagram-illustrating crosswalk between channels in a gigabit channel.

Figure 4 is a block diagram illustrating a system overview of a communications channel.

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Figure 5 is a circuit diagram illustrating a transmitter.

Figure 6 is a graph illustrating a transmit signal.

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Figure 7 is a graph illustrating a composite signal with echoes.

5   Figure 8 is a circuit diagram illustrating a replica transmitter.

Figure 9 is a graph illustrating a receive signal.

10   Figure 10 is block diagram illustrating a low-pass filter.

Figure 11 is a circuit diagram illustrating an active resistive summer.

15   Figure 12 is a circuit diagram illustrating an error detection circuit.

Figure 13 is a circuit diagram illustrating a low-pass filter.

20

Figure 14 is a circuit diagram illustrating a conventional voltage controlled current source.

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DETAILED DESCRIPTION OF THE
PRESENTLY PREFERRED EMBODIMENTS

The preferred embodiments will be described with respect to

5   a gigabit channel, as used, for example, in an Ethernet

network; and to electrical circuits associated with

separating transmit and receive signals in such a gigabit

channel. The preferred embodiments will also be described

with respect to baseline wander correction in such a

10 gigabit channel. However, as will be appreciated by those

skilled in the art, the present invention is also

applicable to other transmission channels, and to other

electrical circuits having applications requiring

cancellation of transmit signals, for example.

15

Figure 4 is a block diagram illustrating principle

components for one of the four channels in a preferred

gigabit channel configuration for use in an Ethernet

network. As illustrated in Figure 4, a vertical dashed

20 line divides analog and digital processing components. The

analog components preferably include a transmitter ("XMTR")

1, replica transmitter ("Replica XMTR") 2, transmit

canceller 3, baseline correction module 4, low pass filter

("LPF") 5, analog-to-digital converter ("ADC") 6, and

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phase-lock loop ("PLL") 7.   A known PLL can be used with

the present invention.

Digital processing components preferably include a

5   transmitter encoder 10, echo module 11, NEXT cancellers 12-

14 to assist in removing echoes, synchronization module 15,

FIR (Finite Impulse Response) equalizer 16 and a DFE

(Decision Feedback Equalizer) 17 to equalize a receive

signal, and a Viterbi module 18.   The digital processing

10   components also include baseline correction modules 19 and

20 to correct residual baseline wander.   A timing recovery

module 21, an error correction detector 22 (described in

further detail below), and summing junction 23 are also

shown.   The individual digital components designated by

15   blocks in Figure 3 are all well known in the communication

arts, and their specific construction and operation are not

critical to the operation or best mode for carrying out the

present invention.

20   The analog "front-end" components shown in Figure 4 will

now be described in even further detail.   The front-end

analog components are preferably designed and constructed

via customized integrated circuits.   However, as will be

appreciated by those skilled in the art, the inventive

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circuits and corresponding configuration could also be
realized using discrete components as well.

    As illustrated in Figure 5, transmitter 1 preferably
5   includes a current-source I_{tx} that generates a transmit
signal over a resistor R_{tx}. An appropriated value for
resistor R_{tx} can be selected to match the line impedance,
for example. In one preferred embodiment, a resistor
center tap is set to 2.5 volts so the transmitter 1
10 effectively sees a differential impedance of 25 ohms.
Preferred performance specifications for the transmitter 1
are further detailed in Table 1, below.

    An impulse transmit signal can be generated from a unit
15 square pulse of 1T width filtered by a one-pole, low-pass
filter (not shown) with a cutoff frequency between 85MHz
and 125MHz. Slew-rate control can also be used to limit
the rise and fall times and thus reduce the high frequency
components of a transmit signal. Of course, any transmit
20 signal preferably fits into the transmit template provided
by the IEEE 802.3ab Standard. An ideal transmit pulse is
shown in Figure 6.

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A measured voltage V_{tx} across R_{tx} (Figure 5) is shown in
Figure 7. The measured signal V_{tx} contains interference
caused by line reflections (e.g., echoes). The reflections
are caused by impedance discontinuity due to impedance

5   mismatch between different cables. For example, a large
reflection pulse at 60ns as shown in Figure 7 corresponds
to a reflection from the impedance discontinuity at an
adapter connecting a 5m patch cord to a 100m cable. The
magnitude of the echoes can be significant when compared to

10   the magnitude of the receive signal at a long line length,
and therefore, echo cancellation, as provided by the NEXT
cancellers 12-14 shown in Figure 4, is employed.

A receive signal V_{rcv} (e.g., a signal received from a far-end

15   transceiver) is also measured across resistor R_{tx}, as shown
in Figure 5. Accordingly, the near end transmit signal
($I_{tx}R_{tx}$) is preferably canceled or reduced from the composite
signal V_{tx} in order to effectively recover the far-end
received signal V_{rcv}. This type of active cancellation can

20   be accomplished with a replica transmit signal V_{txr}.
Accordingly, a replica transmitter 2 (to be described
below) is provided to generate a signal V_{txr} to be
subtracted from the measured signal V_{tx}, thus, effectively
reducing the transmit signal ($I_{tx}R_{tx}$).

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A receive signal x(t) transmitted with pulse amplitude

modulation ("PAM") is define by:

5               $$x(t) = \sum_{n=1}^{\infty} a_n p(t - nT),$$ (7)

where a_n is the transmit symbols and $p(t)$ is the channel

pulse derived by convoluting an impulse transmit pulse with

10   a channel response defined by Equation 5. The receive

signal for a 100m cable is heavily attenuated by the

transmission line and the pulse width is dispersed, as

shown in Figure 9. A 100m UTP delays the signal by about

550ns. Signal equalization preferably uses high frequency

15   boosting via the FIR 16 to remove precursor intersymbol

interference ("ISI") and to insert a zero crossing for

timing recovery 21. The DFE 17 is used to remove

postcursor ISI.

20   The receive signal's elongated tail results from

transformer coupling (e.g., a high-pass filter) with a time

constant (e.g., L/R) typically on the order of micro-

seconds. Since the receive signal contains little or no

average DC energy, the negative tail has the same amount of

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energy as the positive pulse. In this regard, the signal's

area integral is zero. In a typical example, a tail can

last over 10μs with a magnitude of no more than 0.5mV. The

long tail causes any DC bias to drift back toward zero,

5   which can lead to baseline wander. As will be appreciated,

this response time is too long to be practically removed by

a digital equalizer, but the response is slow enough to be

cancelled using a slow integrator, for example. The

baseline wander canceller 4 is preferably decision directed

10 to minimize the error defined by the difference between the

equalized value and it's sliced value, as discussed below.

As illustrated in Figure 8, the replica transmitter 2

includes a current source I_{txr}. I_{txr} is coupled to a voltage

15 V through resistors R, as shown in Figure 8. In a

preferred embodiment, R is 100 ohms and V is about 2.5

volts. The replica signal V_{txr} is preferably filtered

through a known low-pass filter to obtain a low-pass

replica signal ("V_{txrl}"), as shown in Figure 10. Replica

20 signal V_{txr} can also be inverted in a known manner to produce

—V_{txr}. The preferred performance specifications for the

transmitter 1 and replica transmitter 2 are shown in Table

1.

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Table 1: Transmitter and Replica Performance Specifications

Parameters	Specifications
Transmit Current	+/- 40mA
Replica Transmit Current	¼ of transmit current
Number of levels	16 (not including 0)
Number of sub-units	8 (sequentially delayed)
Transmit Profile	[1 1 2 2 1 1], w/ ~1ns delay
Replica Transmit Profile	[1 1 3 3], w/ ~1ns delay
R_{tx}	100Ω

5   A transmit signal canceller 4 is illustrated in Figure 11.

The transmit canceller 4 removes the transmission signal

($I_{tx}R_{tx}$) from the measured (or detected) transmit V_{tx} signal.

In particular, the transmit canceller includes an active

10 resistive summer that provides a large input dynamic range

and stable linearity characteristics, while removing (e.g.,

reducing or canceling) the unwanted transmit signal

component.

15 As illustrated in Figure 11, the active summer includes an

operational amplifier ("op-amp") with inverting feedback.

The op-amp is preferably constructed using integrated

circuits in a known manner. The summer receives V_{txr1}, V_{tx},

$-V_{txr}$, I_{cms}, and I_{bl} as input signals. I_{bl} is a baseline

20 wander control current, and I_{cms} is a common-mode shift

current, each as further discussed below.

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As will be appreciated by those skilled in the art, a
transformer typically has high-pass characteristics.
Accordingly, replica signal $-V_{txr}$ is combined (e.g.,
subtracted via the active resistive summer) with the low

5   pass replica signal V_{txrl} to produce a high-pass replica
signal. As an alternative configuration, V_{txr} could be
filtered through a known high-pass filter prior to the
transmit canceller 3 stage.

10   Returning to Figure 11, receive signal V_{rcv} is determined
from the following relationships.

Let:

15   Vi = voltage for the op-amp's positive terminal;

$V_1{'}$ = V_{txrl};

V_2 = V_{tx};

$-V_3$ = $-V_{txr}$;

i_4 = I_{cmt}; and

20   i_5 = I_{bl}.

Then:

(i) $i_1 + i_2 - i_3 - i_4 - i_5 = i_0$; and

(ii) $\dfrac{V_1 - Vi}{R_1} = i_1$; $\dfrac{V_2 - Vi}{R_1} = i_2$; $\dfrac{Vi - -V_3}{R_1} = i_3$; $\dfrac{Vi - Vrcv}{R_r} = i_0$.

$\Rightarrow$ $\dfrac{V_1 - Vi}{R_1} + \dfrac{V_2 - Vi}{R_1} - \dfrac{Vi + V_3}{R_1} - i_4 - i_5 = \dfrac{Vi - Vrcv}{R_r}$

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$$\Rightarrow \quad \frac{V_1 + V_2 - V_3 - 3Vi}{R_1} - i_1 - i_s = \frac{Vi - Vrcv}{R_r}$$

$$\Rightarrow \quad \frac{R_r}{R_1}(V_1 + V_2 - V_3 - 3Vi) - i_1 \cdot R_r - i_s \cdot R_r = Vi - Vrcv$$

$$\Rightarrow \quad \frac{R_r}{R_1}(V_1 + V_2 - V_3 - 3vi) - R_r i_1 - R_r i_s - Vi = -Vrcv$$

$$V_{rcv} = V_1 - \frac{R_r}{R_1}(V_1 + V_2 - V_3 - 3Vi) + R_r(i_1 + i_s) \tag{8}$$

5

Substituting the input signals for their placeholders
yields the definition for V_{rcv}, as follows:

$$V_{rcv} = Vi - \frac{R_r}{R_1}(Vtxrl + Vtx - Vtxr - 3Vi) + R_r(Icms + Ibl). \tag{9}$$

10

The gain is preferably set between 0.75 and 1 (e.g., R_F/R_1
equals 0.75 to 1). For a small signal analysis, Vi can be
set to zero (0). Also, as will be appreciated by those
skilled in the art, in a fully differential circuit, Vi

15   effectively drops out of the equations since $V_{rcv} = V_{rcv}^{(+)} -$
$V_{rcv}^{(-)}$. As discussed, V_{txrl} and $-V_{txr}$ are combined through the
active summer to provide a high-pass replica signal
("V_{txrh}"). The receive signal V_{rcv} can then be recovered as
shown by Equation 9.

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Preferred transmit canceller specifications are detailed in Table 2, below.

Table 2:　Transmit Canceller Performance Specifications

Parameters	Specifications
Input Dynamic Range	+/-2.5V(diff.) for transmit signal
Output Dynamic Range	+/-1V(diff.)
Input impedance	High, ~10k.
Output impedance	Low
Cutoff frequency	Greater than 31.5 Mhz
DC Gain	0.85 -dependent on the LPF 5 and ADC 6 characteristics (Figure 4)
Power	25 mw, including LPF 5 (Figure 4)
R_f	8.5 KΩ; or 7.5KΩ for increased attenuation
Vi	2.0 volts
R_i	10KΩ

A known current mode circuit, e.g., a voltage controlled current source (VCCS) as shown in Figure 14, with feedback preferably sets the summer input current-mode voltage (V_{cm}). Of course, other known current mode circuits could be employed with the present invention. This current-mode circuit shifts the common-mode of both the transmit and replica transmit signals. The input to the op amp (V_{aip}, V_{ain}) is compared against the desired op amp output common-mode voltage (V_d):

$$V_d = (V_{aip} - V_{cm}) + (V_{ain} - V_{cm}).$$
(10)

Then, the common-mode shift current can be determined from:

$$I_{cms} = V_d g_m + I_0, \qquad\qquad (11)$$

5

where g_m is a transconductance and I_0 is an offset current.
An appropriate transconductance and offset current can be
selected by setting $V_{cm} = I_{cms}R_F = V_d g_m R_F + I_0 R_F$, to ensure a
proper common-mode voltage seen by the op amp inputs.   In
10   this manner, the common mode shift current I_{cms} can be
regulated to pull down the common mode voltage of the
operational amplifier as needed.

Baseline wander current I_{bl} is also "summed" by the active
15   resistive summer, as shown in Figure 11, to correct
baseline wander.   Approximately ninety percent (90%) of all
system baseline correction can be obtained through the
active summer.   The remaining baseline residual can be
digitally corrected through an equalizer, for example.   As
20   will be appreciated, the Figure 11 topology allows the
current sources (I_{bl} and I_{cms}) to each have a fixed output
voltage, thus, minimizing current deviation due to finite
output resistance.

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The baseline wander correction module 4 preferably corrects

for baseline wander using a decision-directed method, such

as a discrete integrator. The decision-directed method can

be implemented with a known charge pump, where the pump

5   sign (e.g., +1/-1) is determined digitally using an error

between the equalized baseline signal (y_k) and a sliced

baseline signal ($\hat{y}_k$), as shown in Figure 12. As will be

appreciated by those skilled in the art, the expected error

value (e.g., $E[e_k]$) is ideally driven to zero. The charge

10   pump is preferably pumped up or down based on the error

value. For example, a positive error implies that a

negative value should be input into the charge pump. For a

negative error, a positive value should be input into the

charge pump. The charge pump preferably has at least two

15   current settings to regulate I_{bl}. Of course, a charge pump

with many current settings could be used to obtain finer

baseline correction control.

The preferred baseline wander correction performance

20   specifications are further detailed in Table 3, below.

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Table 3: Baseline Wander Correction Specification

Parameters	Specifications
Output Dynamic Range	+/-100uA (diff.), (+/-1V/R_1, $R_1 = 10k\Omega$)
Output impedance	High
Integration Factors	2 mV/T, 4 mV/T
Bandwidth	> 100MHz

A second-order low-pass filter, as shown in Figure 13, is

5   cascaded after the summer to preferably flatten the

frequency response out to about 31.25MHz (<1dB). A minimum

overall attenuation of 20dB at 125MHz is desirable for the

low pass filter. In a sampled system, some aliasing beyond

Nyquist frequency (or excess bandwidth) is acceptable, but

10   minimum aliasing is allowed at the sampling frequency. The

transmitted data is preferably band-limited to the Nyquist

rate.

Preferred performance characteristics of the low pass

15   filter 5 are further detailed in Table 4, below.

Table 4: LPF Performance Specification

Parameters	Specifications
Input Dynamic Range	+/-1V(diff.)
Output Dynamic Range	+/-1V(diff.)
Input impedance	High, ~10k.
Output impedance	Low
Cutoff frequency	50-60Mhz.
Q (2nd order)	~1
Input impedance	High, ~10k.
Output impedance	Low, <100
DC gain	1

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As an alternative arrangement, a third-order Sallen and Key low pass filter as disclosed in a co-pending application by the same inventor of this application, titled "CALIBRATION

5   CIRCUIT," filed concurrently herewith, and hereby incorporated by reference, could be used as filter 5. Similarly, the calibration circuit disclosed therein could also be used to calibrate the low pass filter 5.

10   Analog-to-digital converters are well know in the art. As will be appreciated, the ADC 6 resolution is often determined by system digital processing requirements. In a preferred embodiment, the Viterbi detector 18 requires an effective 7-bit resolution. Residual baseline wander,

15   echoes, and crosstalk increase the dynamic range by about 200-300mV, which increases the required resolution. The reduction in dynamic range due to insertion loss for a 100m cable is approximately 40%. Accordingly, an 8-bit resolution is preferred.

20

The preferred ADC performance specifications are further detailed in Table 5, below.

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Table 5: ADC Performance Specification

Parameters	Specifications
Resolution	8-bits minimum.
Sampling frequency	125MS/s
Source Output Impedance	Low, ~200-400Ω

5   Thus, a transmit canceller including an active resistive

summer has been described. Such an active resistive summer

has not heretofore been developed for applications such as

canceling signals in gigabit channels. Correcting baseline

wander through such an active resistive summer has also

10   been described herein.

While the present invention has been described with respect

to what is presently considered to be the preferred

embodiments, it will be understood that the invention is

15   not limited to the disclosed embodiments. To the contrary,

the invention covers various modifications and equivalent

arrangements included within the spirit and scope of the

appended claims. The scope of the following claims is to

be accorded the broadest interpretation so as to encompass

20   all such modifications and equivalent structures and

functions.

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For example, while preferred circuit configurations and component values have been described, it will be understood that modifications could be made without deviating from the inventive structures. For example, values for the feedback

5   and input resistors R_f and R_1 could be changed to obtain higher or lower gains. Also, an active resistive summer could be constructed to sum only the measured signal V_{tx} and the replica signal V_{txr} (or a high-pass version of the replica), for example. Additionally, while the

10 communication channel has been described with respect to a twisted pair of cables, the invention may also be practiced with other communication channels such as optical and wireless channels. Moreover, this invention should not be limited to gigabit transmission rates and can be practiced

15 at any transmission rate requiring the signal processing characteristics of the invention. Of course, these and other such modifications are covered by the present invention.